Design of a power-line modem for monitoring and control systems

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Abstract

We design a power-line communication (PLC) modem applicable to central monitoring and control systems. To make the designed modem robust to frequency-selective and time-varying channel condition, we consider the use of a multicarrier CPFSK modulation scheme. We employ an adaptive impedance loading scheme to reduce the power loss due to impedance mismatch between the channel and analog front end of the modem. The performance of the designed modem is verified by computer simulation. Numerical results show that the use of the proposed modem can transmit data reliably with small power loss in frequency selective channel.

1. Introduction

There have been increasing demands for central control systems for automation of commercial buildings. Since the power-line network is ubiquitous and inexpensive, the nature of power-line may not be appropriate for reliable communications [1]. In particular, the power line communication (PLC) has been considered as one of the major schemes for building automation [2]. It is well understood that the power line has unpredictable and frequency-selective characteristics. The gain variation can be very large depending upon the position and kind of the load [1]. The gain variation due to capacitive loads can be as much as -100dB/km. The variation of the input impedance of the PLC channel can cause a serious injection power loss due to the impedance mismatch [1, 2]. The nature of power-line may not be appropriate for reliable communications [1].

Early PLC modems have been developed for control systems using a modulation scheme with low spectral density such as the FSK, spread spectrum (SS) and binary phase shift keying (BPSK)[1, 3, 4]. Since the SS scheme requires large bandwidth to obtain a large processing gain, it may not be appropriate for spectral efficient transmission [5]. Although Echelon considers the use of BPSK PLC modem for building control systems [2], it may not be practical to employ coherent detection under harsh and unpredictable power-line channel condition. Moreover, these previous works did not consider frequency-selectivity and impedance mismatch problems in the PLC channel.

To mitigate such hostile power-line channel condition, we consider to design a transceiver scheme that employs an adaptive technique for impedance loading, with the use of a proper MAC protocol for central control. The carrier sensing multiple access-collision avoidance (CSMA-CA) method can provide high throughput efficiency, but it may not be able to reliably detect collisions in the PLC environment [6]. To alleviate this problem, we consider the use of a polling protocol based on a dedicated reservation method to reliably detect the frame timing under severe PLC condition.

The use of frequency diversity techniques has widely been considered to obtain reliable transmission performance in frequency selective channels [7, 8]. We consider the use of a multicarrier FSK scheme to obtain the frequency diversity effect, where the maximal ratio combining (MRC) or equal gain combining (EGC) method is considered.

The impedance of the power-line channel varies abruptly due to the change of electrical loads. The impedance mismatch of the analog front end to the power-line can cause high injection loss of the transmits power. It was reported that the impedance can be adjusted using a compensator circuit comprising passive components [9]. Since this method requires an additional ASIC chip, it may not be suitable for simple control systems. We propose an impedance estimation scheme utilizing the voltage of line coupler without the use of an additional pilot symbol. Since the voltage of the line coupler is a function of the source to the channel impedance ratio, it can be used for adjustment of the impedance of the transmitter.

This paper is organized as follows. The structure of the designed PLC modem and MAC protocol is described in

Section II. Section III describes a multicarrier FSK scheme for frequency diversity gain. A noble adaptive impedance matching algorithm is described and verified in Section IV, and conclusions are drawn in Section V.

2. System structure

Fig 1. depicts the structure of the designed PLC modem for centralized control systems. The proposed modem employs a binary CPFSK scheme with three carriers to reliably transmit the data reliably in frequency selective PLC channel. We consider the use of a frequency range of 100 to 450kHz considering the noise and interference in lower frequency range. For ease of implementation, we choose the three frequencies (150, 250 and 350kHz), as the carrier frequency. An equal gain combining scheme with non-coherent detection is employed for frequency diversity and the use of a Reed-Solomon code is considered to make it robust to impulse noise. We use a Reed-Solomon code with $(n,k,m)=(7,5,3)$ considering the length of impulse noise.

A MAC protocol is required to manage the channel allocation/re-allocation between the slave devices. In a centralized control system, it is required to employ a MAC protocol that can provide reliable control channel connection between the control center and slave modems. Although random access protocols such as the CSMA have high throughput efficiency, they may not be able to reliably detect the collisions due to the characteristics of the PLC channel. In addition, the random access protocol requires accurate timing recovery for each received slot.

To overcome these problems, we employ a polling protocol with dedicated reservation to have robust control channel connection, where each slave modem has its unique identification (ID) number. During the initialization process, the master modem transmits a control slot to the first slave modem and waits a response message from the first one and then it moves to the second one.

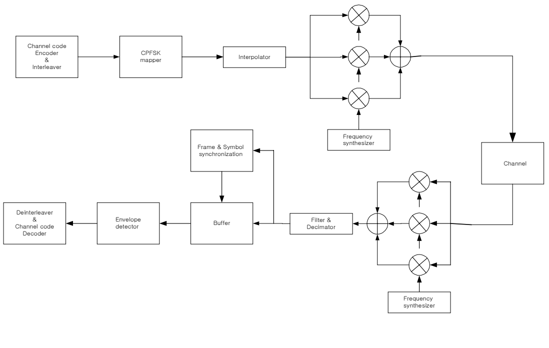


Fig 1. Block diagram of the designed modem

If the master modem fails to receive the response message from the slave modem, it resends the initialization message to the slave modem. If the master modem fails to receive the response from slave once again, it sends the initialization message to the next slave modem. In this manner, the master modem sends the initialization message to all the slave modems and can make a reservation with the ID number based on the response of the salve modems.

We use a special training slot as depicted in Fig. 2 for synchronization during the initialization process. After the synchronization, a normal slot is transmitted to the slave modems for device control. A pseudo-noise (PN) sequence is repeatedly inserted into the training slot, repeatedly as depicted in Fig. 2. (a). Timing recovery can be achieved using correlation characteristics of these PN codes. The normal slot includes the information bits for the control message, parity check bit for ARQ, RS code and PN sequence for timing recovery. Using this slot format, a total 127 number of devices can be controlled.

3. Frequency diversity

The combining technique is an effective method for combating the frequency-selectivity problem in a noncoherent FSK scheme. Although the maximum ratio combining (MRC) can provide better performance of envelope detector, it requires the channel information. For simplicity of implementation, we consider the use of equal gain combining (EGC) that does not need channel information. It can be approximated that the PLC channel amplitude normalized with respect to the time averaged value is statistically Rician distributed [10]. The performance of noncoherent FSK modulation was analyzed in Nakagami fading channel using the characteristic function method [7]. In this paper, we analyze the error probability of a binary FSK transceiver with EGC in PLC channel.

PN code (13 bits)	PN code (13 bits)	RSVD (13 bits)

(a) Training slot

PN code (13 bits)	Address (7 bits)	Control message (7 bits)	Parity check (1bit)	Rs code redundancy (6 bits)

1.18 msec

(b) Normal slot

Fig 2. Slot format of the designed modem

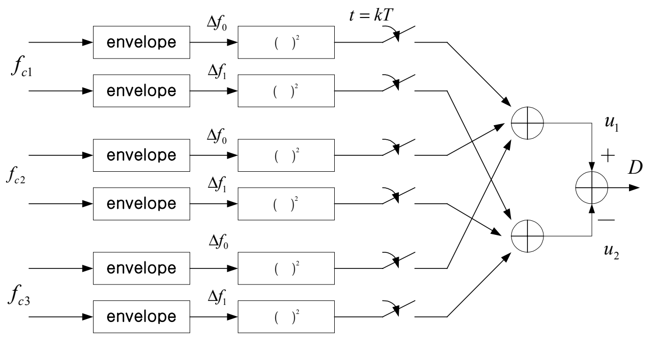


Fig 3. Postdetection equal gain combiner for noncoherent FSK

Consider L diversity branches over a flat Rician fading channel as depicted in Fig. 3, where $L = 3$. The received signal of the k-th branch is given by

$$r_k(t) = \mathrm{Re}\left\{\left|\alpha_k e^{-j\phi_k} s(t) + n_k(t)\right| e^{j2\pi f_c t}\right\}, \quad k = 1,\cdots,L \quad (1)$$

where $s(t)$ denotes the baseband signal, α_k is the amplitude and ϕ_k is the phase of the k-th diversity branch, f_c is the carrier frequency and $n_k(t)$ denotes additive Gaussian noise. Representing the branch gain vector as

$$\left[\alpha_1 e^{-j\phi_1},\ldots,\alpha_L e^{-j\phi_L}\right]^T = \mathbf{X}_c + j\mathbf{X}_s \quad (2)$$

the covariance matrix $\mathbf{K}$ and mean vector of this channel gain can be written as

$$\mathbf{K} \triangleq \begin{bmatrix} \mathbf{K}_{cc} & \mathbf{K}_{cs} \\ \mathbf{K}_{cs}^T & \mathbf{K}_{ss} \end{bmatrix}, \quad \bar{\mu} \triangleq \begin{bmatrix} \bar{\mu}_c \\ \bar{\mu}_s \end{bmatrix} \quad (3)$$

where the superscript 'T' denotes the transpose of a vector, $\mathbf{K}_{cc} = E\{\mathbf{X}_c\mathbf{X}_c\}$, $\mathbf{K}_{cs} = E\{\mathbf{X}_c\mathbf{X}_s\}$, $\mathbf{K}_{ss} = E\{\mathbf{X}_s\mathbf{X}_s\}$, $\bar{\mu}_c = E\{\mathbf{X}_c\}$ and $\bar{\mu}_s = E\{\mathbf{X}_s\}$. Here $E\{\mathbf{X}\}$ denotes the expectation of $\mathbf{X}$. Let us define $\bar{\beta}$ by

$$\bar{\beta} \triangleq [\beta_1, \beta_2, \ldots, \beta_L]^T = \left[\alpha_1^2, \alpha_2^2, \ldots, \alpha_L^2\right]^T \quad (4)$$

To analyze the error performance, we need the characteristic function of $\beta_{tot} = \sum_{k=1}^{L} \beta_k$. It can be shown that $\mathbf{K}$ can be represented as

$$\mathbf{K} = \mathbf{R}\Lambda\mathbf{R}^T \quad (5)$$

where $\Lambda = \mathrm{diag}(\lambda_1,\ldots,\lambda_{2L})$ is a diagonal matrix of comprising eigenvalue of $\mathbf{K}$ and $\mathbf{R}$ is a unitary matrix. The characteristic function of β_{tot} can be obtained [8]

$$\Psi_{\beta_{tot}}(j\omega) = \frac{\exp\left\{ j\omega \sum_{k=1}^{2L} \frac{\eta_k^2}{(1-2j\omega\lambda_k)} \right\}}{\prod_{k=1}^{2L} (1-2j\omega\lambda_k)^{1/2}} \quad (6)$$

where $\bar{\eta} = \mathbf{R}^T \bar{\mu}$. The transmitted baseband signal $s(t)$ for binary data symbol i, is given by

$$s(t) = \sqrt{\frac{2E_s}{T_s}} e^{j2\pi(\Delta f_i)t}, \quad (7)$$

where E_s is the energy per symbol, T_s is the symbol duration, $\Delta f_0 = -\Delta f$ and $\Delta f_1 = \Delta f$. As depicted in Fig. 3, the decision variable D after the EGC is given by

$$D = \sum_{k=1}^{L} u_{k1} - \sum_{k=1}^{L} u_{k0} \quad (8)$$

where

$$u_{ki} = \left|\{s(t) * h_k(t)\} \cdot e^{j(2\pi\Delta f_i + \theta)}\right|^2 \quad (9)$$

Here, $h_k(t)$ denotes the channel impulse response of k-th branch and θ is the phase offset.

Let γ be the signal to noise power ratio (SNR) at the combiner's output. Then, we have

$$\gamma = \sum_{k=1}^{L} \gamma_k = \frac{E_s}{N_0} \beta_{tot} \quad (10)$$

where $\gamma_k (= E_s \alpha_k^2 / N_0)$ is the instantaneous SNR at k-th branch. For a given output SNR γ, the conditional charateristic function of D can be obtained as

$$\Psi_D(j\omega \mid \gamma) = \frac{\exp\left\{ \frac{4j\omega E_s N_0}{(1-4j\omega E_s N_0)} \gamma \right\}}{(1+4j\omega E_s N_0)^L (1-4j\omega E_s N_0)^L} \quad (11)$$

By averaging $\Psi_D(j\omega)$ over the probability density function (pdf) of γ and using (6), the characteristic function of D is represented as

$$\Psi_D(j\omega) = \frac{\exp\left\{ 4j\omega E_s N_0 \sum_{k=1}^{2L} \frac{b_k}{1-4j\omega E_s N_0 [1+a_k]} \right\}}{(1+4j\omega E_s N_0)^L \prod_{k=1}^{2L} (1-4j\omega E_s N_0 [1+a_k])^{1/2}} \quad (12)$$

where $a_k = 2\lambda_k E_s / N_0$, $b_k = \eta_k^2 E_s / N_0$.

Using the inversion theorem [11], the BER can be calculated as

$$P_e = -\frac{1}{(L-1)!} G^{(L-1)}(z) \Big|_{z=-1} \quad (13)$$

where

$$G(z) \triangleq (1+z)^L \frac{\Psi_D(z/(4E_s N_0))}{z} \quad (14)$$

It can be shown that using Faa di Bruno's formula [12] that

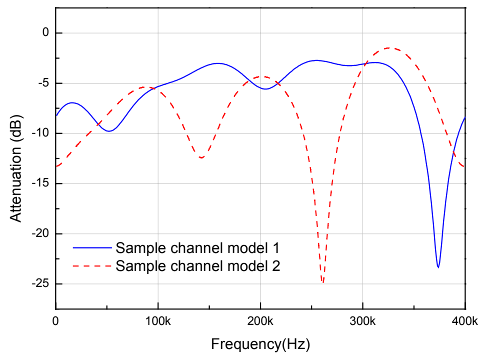
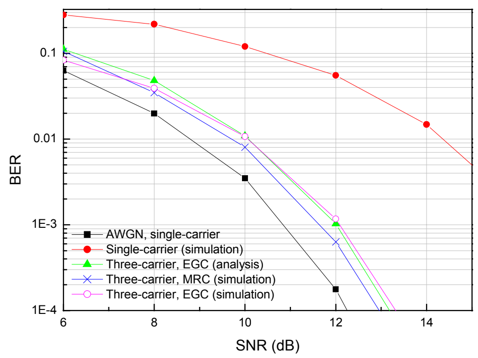
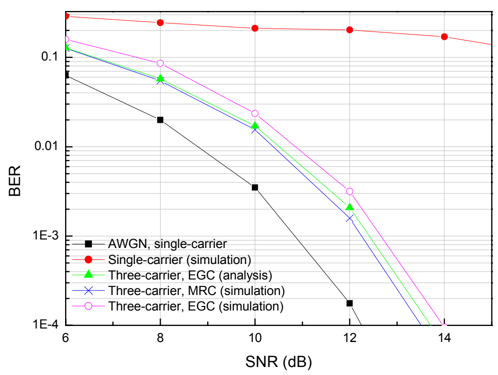


Fig. 4 Frequency response of sample PLC channel model



(a) Sample channel model 1



(a) Sample channel model 2

Fig. 5 BER performance of the EGC scheme over sample PLC channel

$$P_e = \frac{\exp\left\{-\frac{E_s}{N_0}\sum_{k=1}^{2L}\frac{\eta_k^2}{2+2\lambda_k E_s/N_0}\right\}}{\prod_{k=1}^{2L}\left(2+2\lambda_k E_s/N_0\right)^{1/2}}$$

$$\times \sum_{\substack{(l_1,l_2,\ldots,l_{L-1}) \\ 0\le l_1,l_2,\ldots,l_{L-1}\le L-1 \\ l_1+2l_2+\cdots+(L-1)l_{L-1}=L-1}} \left(\prod_{m=1}^{L-1}\frac{1}{l_m!}\left[\frac{1}{m}+\frac{1}{2m}\sum_{k=1}^{2L}\frac{\left(1+2\lambda_k E_s/N_0\right)^m}{\left(2+2\lambda_k E_s/N_0\right)^m} \quad (15)\right.\right.$$

$$\left.\left.\frac{E_s}{N_0}\sum_{k=1}^{2L}\eta_k^2\frac{\left(1+2\lambda_k E_s/N_0\right)^{m-1}}{\left(2+2\lambda_k E_s/N_0\right)^{m+1}}\right]^{l_m}\right)$$

where l_i, $i=1,2,\cdots,L-1$, is a nonnegative integer.

The accuracy of BER analysis and the performance of the EGC scheme are verified by computer simulation in Rician PLC channel model in [10]. Fig. 4 illustrates the frequency response of two sample PLC channel models.

The analytic BER performance of the proposed scheme with EGC calculated is verified by computer simulation in Fig. 5., where the performance of the MRC scheme is also shown for comparison. It can be seen that the use of a single-carrier scheme cannot be applied to frequency selective PLC channel as in the sample channel-2 and that the use of a multi-carrier scheme with a simple diversity scheme such as EGC is quietly effective in PLC channel. It can also be seen that the analytic result agrees well with the simulation result.

It can be seen that the both EGC and MRC schemes provide similar performance and that multicarrier scheme is better than the single carrier scheme. It can be seen that the performance of single-carrier scheme degrades severely in sample channel-2 due to frequency selective null and that the BER performance degradation can be compensated by employing the simple EGC scheme. As depicted in Fig. 5, the BER of EGC scheme over the PLC channel can be predicted approximately using analysis result. Since we assume flat fading over one FSK branch bandwidth ($f_{c_i}-\Delta f \sim f_{c_i}+\Delta f$) in analysis, there are some differences between computer simulation result and analysis result.

4. Adaptive impedance loading

In PLC environment, the impedance of the channel can abruptly vary due to the change of electrical loads. This impedance change causes serious impedance mismatch problem in the transmitter. As a result, the coverage range can significantly be reduced by reflective signals due to impedance mismatch and the dynamic range of the received signal increases according to the variant of the load impedance.

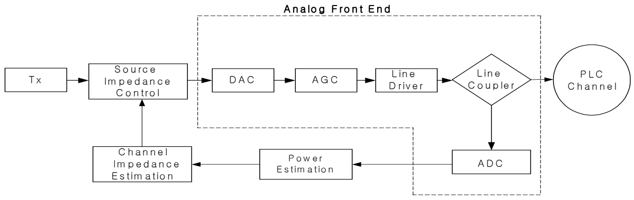


Fig 6. Impedance matching structure

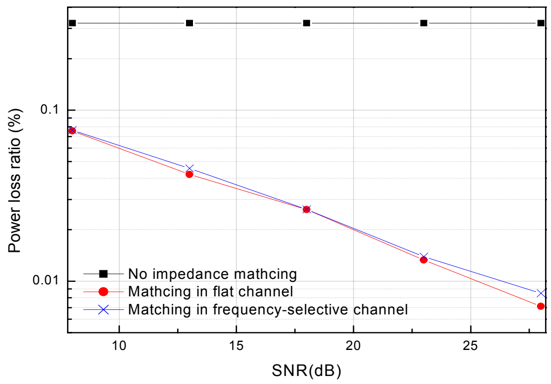


Fig 7. Power delivery efficiency of the proposed impedance loading method.

It is desirable for the transmit modem to match its source impedance Z_S to the channel impedance Z_L. It is necessary to accurately estimate impedance information for impedance matching, which is complex valued due to resistive and reactance components. We consider an algorithm that needs only the magnitude of the channel impedance. To estimate the channel impedance, Fig 6 depicts the proposed impedance matching block including the analog front end (AFE).

Representing the complex valued channel impedance as

$$Z_L = Z_{L_r} + iZ_{L_i} \tag{16}$$

the output of the transmitter is given by

$$|V_L|^2 = \frac{{Z_{L_r}}^2 + {Z_{L_i}}^2}{\left(Z_S + Z_{L_r}\right)^2 + {Z_{L_i}}^2}|V_S|^2 \tag{17}$$

where V_S denotes the source voltage of the transmitter.

The magnitude of the channel impedance can be estimated as

$$|Z_L|^2 = \frac{\xi^2 {Z_S}^2}{1-\xi^2}\left(1+\frac{2Z_{L_r}}{Z_S}\right) \tag{18}$$

where $\xi = |V_L|/|V_S|$. Since the real part Z_{L_r} of the channel impedance is unknown, it is necessary to know Z_{L_r}/Z_S. Assuming that the source impedance Z_S of the transmitter is very large compared to Z_{L_r}, the channel impedance can approximately be estimated as

$$|Z_L|^2 \cong \frac{\xi^2 {Z_S}^2}{1-\xi^2} \tag{19}$$

To evaluate the efficiency of the proposed loading algorithm, Fig 7 depicts the power delivery efficiency in terms of the SNR. It can be seen that the power loss is nearly flat regardless of the SNR without impedance matching and that the power loss is reduced with the use of the proposed scheme because the channel impedance can be estimated more accurately as the SNR increases.

5. Conclusion

We have considered the design of power-line modem for central control of building automation. To mitigate hostile channel condition, we proposed the use of a multicarrier FSK scheme with adaptive impedance matching. Numerical results show that the proposed scheme is quite applicable to real environments with reduced implementation complexity. Also we can reduce the injection power loss by the use of the simple adaptive impedance loading scheme.

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